An Algorithm for Multichannel Coherent Digital Communications Over Long Range Underwater Acoustic Telemetry Channels

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Abstract - The problem of achieving reliable digital communications over long range underwater acoustic telemetry channels is addressed, and a receiver algorithm for multichannel coherent data detection is presented. The receiver consists of a $T/2$ fractionally spaced bank of feedforward equalizers, a multichannel carrier phase synchronizer and a common decision feedback section of the equalizer. An adaptive algorithm is derived based on joint minimum mean squared error optimization of the receiver parameters. The equalizer tap coefficients and estimates of the carrier phases are updated using a combination of a recursive least squares algorithm and a second order multichannel digital phase locked loop. Since the equalizer accomplishes the function of symbol synchronization, no separate delay locked loops are necessary.

The algorithm is successfully applied to the experimental data. The results assert feasibility of coherently combining multiple arrivals in each of the diversity channels, and demonstrate additional spatial diversity improvement.

I. INTRODUCTION

Although the underwater acoustic (UWA) channel fits into the general description of a rapidly fading channel, many of the classical communication techniques, such as those designed for the VHF/UHF mobile radio channels, are not directly applicable to the UWA channel due to its many unique characteristics. Severe performance degradation encounterd on the UWA channels is largely due to the extended, time varying multipath and phase instabilities, the latter of particular concern for long range telemetry [1]. While typical multipath spreads in the mobile radio channel are two or three symbol intervals, in the long range UWA channel they increase to several tens of symbol intervals for moderate to high data rates. On the other hand, a receiver capable of coherently processing multiple arrivals benefits from both power efficiency of coherent detection and diversity improvement inherent in multipath propagation. Due to the rapid fluctuations of the ocean channel, such a receiver is likely to be computationaly intensive, and is often not considered feasible. Fortunately enough, the computational complexity is not of a major concern in the UWA communications, since the data rates are considerably lower than those used in the majority of existing communication media.

In order to account for both amplitude and phase fluctuations of the UWA channel, we address the problem of jointly adaptive synchronization and channel equalization. Joint estimation procedures are known to yield better results than marginal estimation, and the fundamental work in this area was presented in [2], [3]. Since both of these references concentrate on the optimal, maximum likelihood sequence estimation principles, which become inacceptably complex for long channel responses spanning more than, say, ten symbol intervals, we focus on a suboptimal receiver structure with a decision feedback equalizer (DFE). The concept of joint carrier and symbol synchronization, and DFE [4] is extended here to multichannel, or spatial diversity reception, which provides additional improvement in performance with respect to fading and noise. To meet the rapidly changing conditions in the channel, the receiver operates using a combination of recursive least squares (RLS) algorithm for equalizer coefficients update, and a second order multichannel digital phase locked loop (DPLL) for carrier phase synchronization. The receiver requires only two samples per symbol interval, and is suitable for an all digital implementation.

After considering the general receiver structure in Section II, the derivation of the receiver algorithm is given in Section III. Section IV presents some of the results obtained with the experimental long range UWA telemetry data provided by the Woods Hole Oceanographic Institution (WHOI).

II. COMMUNICATION SYSTEM STRUCTURE

We consider a general class of linear, bandwidth efficient modulation techniques. The transmitted signal is represented in its equivalent complex baseband form as

$$u(t)=\sum_{n} d(n)g(t-nT) \tag{1}$$

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Figure 1: Signaling frame.

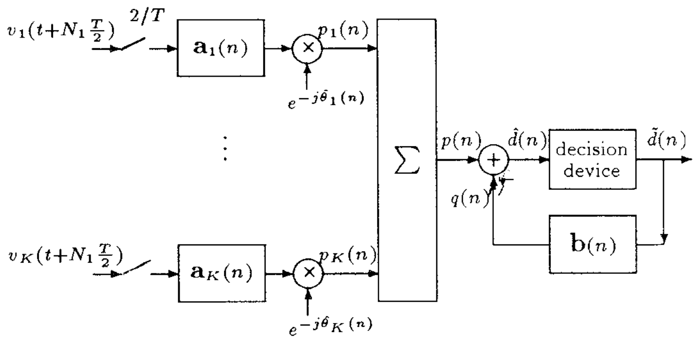

Figure 2: Receiver structure.

where $\{d(n)\}$ is the sequence of independent M-ary data symbols, T is the signaling interval, and $g(t)$ is the basic transmitter pulse usually bandlimited to $1/T$. The signal from the transmitter is sent over K diversity channels, which are assumed to be independently fading. The received signal in each of the channels is quadrature downconverted, lowpass filtered, and A/D converted. Prior to processing, signals from different channels are frame synchronized. This is accomplished using filters matched to the channel probe. The signaling format, containing the channel probe and the data block, is shown in Fig.1. After coarse alignment in time, the received signal in the i^{th} diversity branch is represented as

$$v_i(t) = \sum_n d(n)h_i(t-nT,t)e^{j\theta_i(t)} + \nu_i(t) \qquad (2)$$

where $h_i(\tau,t)$ is the channel impulse response (including any transmit or receive filtering) as a function of delay τ at time t, $\theta_i(t)$ is the carrier phase, and $\nu_i(t)$ is assumed to be white Gaussian noise.

The receiver structure is shown in Fig.2. The front section of the receiver consists of a bank of feedforward equalizers and a multichannel carrier synchronizer. Since the channel impulse responses $h_i(t)$ are not known, there is no explicit matched filtering prior to equalization. The feedforward linear equalizers are fractionally spaced (FS), i.e. they operate on the sequence of input samples taken at time intervals less than T. An infinitely long feedforward equalizer with optimal tap setting accomplishes both functions of matched filtering (adaptive, in case of time varying channels), and optimal T-spaced linear equalization [5]. The most important feature of a FS feedforward section is that it is insensitive to the timing phase of the incomming signals, therefore eliminating the need for separate esimation of the symbol timing. Only if very long, or continuous messages are being transmitted, the need may arise for some sort of adaptive adjustment of the timing phase, in order to ensure the correct position of the center tap of the equalizer [6]. Without loss of generality, we use a fractional spacing of $T/2$, which is adequate for any signal bandwidth less than $1/T$.

Although theoretically the optimally chosen complex tap weights of the linear equalizer correct for any phase offset in the received signal, this is not the case in practice. The carrier phase in the i^{th} channel, $\theta_i(t)$, can be modeled as a sum of three terms: constant phase offset, Doppler frequency shift, and random phase jitter. While an adaptive equalizer is capable of correcting for the constant phase offset and possibly some slow variations of the carrier phase, the carrier frequency offset, as well as more rapid phase fluctuations, result in the equalizer tap rotation phenomenon [7]. This increases the misadjustment noise, and may eventually cause the equalizer taps to diverge. Typically, the tap gains should not change by more than few percent from one symbol interval to another [8]. Therefore, the addition of a carrier phase synchronization loop is necessary to ensure proper operation of the equalizer, especially in the conditions of large phase fluctuations encountered in the UWA channels.

Due to the fact that the feedforward equalizer introduces a delay of a certain number N_1 of symbol intervals, the current estimate of the carrier phase, $\hat{\theta}_i(n)$, lags behind the true phase of the input signal, $\theta_i(nT+N_1T/2)$. It is for this reason that the carrier phase synchronization is performed after equalization, now termed passband equalization, thus eliminating the problem of delay in the phase estimate [7]. The feedforward equalizer output is produced once per symbol interval, and the carrier phase update is performed accordingly. Depending on the particular channel characteristics, it may not be necessary to have a separate DPLL for each of the diversity branches. In the application of interest, however, we found that due to the possibly large differences in time varying Doppler frequency offsets caused by unpredictable motion of the receiver array, it was necessary to have as many phase estimators as diversity channels. This is also one of the reasons that preclude the use of a passband DFE structure [8] in the multichannel form. In this structure, the carrier phase correction is moved further into the decision feedback loop, resulting in some minor improvements.

After coherent combining, the signals from different channels are fed into the common decision feedback part

of the equalizer. Such a structure resembles maximal ratio combining, in which each diversity signal is weighted proportionally to its strength, and coherently combined with the others prior to decision making [5]. Indeed, if there were no intersymbol interference (ISI), the two structures would be equivalent.

Since the channel is time varying, so are the optimal values of receiver parameters. An adaptive algorithm for joint estimation of equalizer coefficients and carrier phases is presented in the following section. The receiver initially operates in the training mode, using known data symbols, after which it is switched to a decision directed mode.

III. DERIVATION OF THE RECEIVER ALGORITHM

Having established the receiver structure, we can proceed to determine the optimal values of its parameters. The optimization criterion we use is the minimum mean squared error (MSE) between the estimated data symbol $\hat{d}(n)$ and the transmitted symbol $d(n)$. The receiver parameters are the tap weights of the multichannel feedforward equalizer, feedback equalizer coefficients, and the carrier phase estimates. In general, there are two ways of computing the equalizer parameters. One is the direct adaptation of the equalizer coefficients driven by the output error, and the other is their computation from the estimated channel impulse response. Although the latter is potentially more robust to the time variations of the channel [9], we chose the usual, direct method, as computationally less involved.

Assuming the constant channel impulse response and carrier phase in some short interval of time, one arrives at the optimal values of equalization and synchronization parameters. Let the i^{th} channel feedforward equalizer tap weight vector be

$$\mathbf{a}_i' = [a^i_{-N_1} \cdots a^i_{N_2}]^* \tag{3}$$

where $(\cdot)'$ denotes the conjugate transpose, and the tap weights are taken as conjugate for later convenience of notation. The input signal samples stored in the i^{th} feedforward equalizer at time nT are conveniently represented in a column vector

$$\mathbf{v}_i(n) = [v(nT + N_1T/2) \cdots v(nT - N_2T/2)]^T. \tag{4}$$

The output of the i^{th} feedforward equalizer, after phase correction by the amount $\hat{\theta}_i$, is given as

$$p_i(n) = \mathbf{a}_i'\mathbf{v}_i(n)e^{-j\hat{\theta}_i} \tag{5}$$

and the coherent combination of all diversity channels is

$$p(n) = \sum_{i=1}^{K} p_i(n). \tag{6}$$

The feedback filter coefficients are arranged as a vector

$$\mathbf{b}' = [b_1 \cdots b_M]^* \tag{7}$$

and the column vector of M previous decisions, currently stored in the feedback filter, is denoted as

$$\tilde{\mathbf{d}}(n) = [\tilde{d}(n-1) \cdots \tilde{d}(n-M)]^T. \tag{8}$$

This defines the output of the feedback filter as

$$q(n) = \mathbf{b}'\tilde{\mathbf{d}}(n). \tag{9}$$

The estimate of the data symbol at time n is

$$\hat{d}(n) = p(n) - q(n) \tag{10}$$

from which the decision $\tilde{d}(n)$ is obtained as the closest signal point. The estimation error is

$$e(n) = d(n) - \hat{d}(n) \tag{11}$$

and the receiver parameters are optimized based on joint minimization of the MSE with respect to $\{\mathbf{a}_i\}, \mathbf{b}$, and $\{\hat{\theta}_i\}$.

In order to find the optimal values of the equalizer coefficients, it is convenient to group all the coefficients into a composite vector $\mathbf{c}$, and to express the estimate $\hat{d}(n)$ as

$$\begin{aligned} \hat{d}(n) &= [\mathbf{a}_1' \cdots \mathbf{a}_K' \; -\mathbf{b}'] \begin{bmatrix} \mathbf{v}_1(n)e^{-j\hat{\theta}_1} \\ \vdots \\ \mathbf{v}_K(n)e^{-j\hat{\theta}_K} \\ \tilde{\mathbf{d}}(n) \end{bmatrix} \\ &= \mathbf{c}'\mathbf{u}(n). \end{aligned} \tag{12}$$

The MSE can now be expressed as a function of the composite equalizer vector $\mathbf{c}$,

$$\begin{aligned} E &= E\{|d(n) - \mathbf{c}'\mathbf{u}(n)|^2\} \\ &= R_{dd} - 2Re\{\mathbf{c}'\mathbf{R}_{ud}\} + \mathbf{c}'\mathbf{R}_{uu}\mathbf{c} \end{aligned} \tag{13}$$

where we have used the notation $\mathbf{R}_{xy} = E\{\mathbf{x}(n)\mathbf{y}'(n)\}$ for the crosscorrelations. The value of $\mathbf{c}$ which minimizes the MSE is the well known solution to the finite order Wiener filtering problem, and is given by

$$\mathbf{c} = \mathbf{R}_{uu}^{-1}\mathbf{R}_{ud}. \tag{14}$$

The optimal values of the estimates of the carrier phases, $\hat{\theta}_i$, are most easily found if the estimate $\hat{d}(n)$ is represented as

$$\begin{aligned} \hat{d}(n) &= p_i(n) + \sum_{j \neq i} p_j(n) - q(n) \\ &= \mathbf{a}_i'\mathbf{v}_i(n)e^{-j\hat{\theta}_i} + f_i(n). \end{aligned} \tag{15}$$

The second term in the last expression is independent of $\hat{\theta}_i$, which makes it possible to express the MSE as

$$\begin{aligned} E &= E\{|d(n) - f_i(n) - \mathbf{a}_i'\mathbf{v}_i(n)e^{-j\hat{\theta}_i}|^2\} \quad (16) \\ &= -2Re\{\mathbf{a}_i' E\{\mathbf{v}_i(n)[d(n) - f_i(n)]^*\}e^{-j\hat{\theta}_i}\} \\ &+ \text{ terms independent of } \hat{\theta}_i \end{aligned}$$

The optimal values $\hat{\theta}_i$ satisfy the gradient equations

$$\frac{\partial E}{\partial \hat{\theta}_i} = -2Im\{\mathbf{a}_i' E\{\mathbf{v}_i(n)[d(n) - f_i(n)]^*\}e^{-j\hat{\theta}_i}\} = 0, \quad (17)$$
$$i = 1, \ldots, K.$$

In order to be able to track the actually time varying optimal solution for the receiver parameters, the equations (14), (17) should be solved recursively, using updated values of possibly time varying crosscorrelations. In the case of a rapidly changing channel, the adaptation has to be carried out continuously. An alternative method to continuous adaptation is the so called block adaptation [10], in which the receiver parameters are updated only during short training blocks interspersed in the data stream, and interpolated between such blocks. The adavantage of such an approach is the prevention of error propagation.

As it was pointed out earlier, the carrier recovery process can theoretically be absorbed in the proces of equalization. It can be verified that the optimal solution in such case would be the same as the one represented by equations (14), (17). The point of having separate expressions for the equalizer coefficients and the carrier phases, is to be able to derive different tracking strategies for the two, which ultimately eliminates the problem of equalizer tap rotation.

The simplest form of an adaptive algorithm is the combination of a least mean squares (LMS) algorithm for the equalizer coefficients updata, and the first order DPLL [7]. Such an algorithm, however, failed on the UWA channel, primarily due to the poor phase tracking capabilities. In order to obtain improved phase tracking capabilities, we introduced a second order DPLL into the process of joint synchronization and equalization. Using the analogy between the phase detector output of a classical DPLL and the instantaneous estimate of the MSE gradient with respect to $\hat{\theta}_i$, we define

$$\Phi_i(n) = Im\{\mathbf{a}_i' E\{\mathbf{v}_i(n)[d(n) - f_i(n)]^*\}e^{-j\hat{\theta}_i}\} \quad (18)$$

as the equivalent output of the i^{th} phase detector. Using the fact that

$$d(n) - f_i(n) = p_i(n) + e(n) \quad (19)$$

the expression (18) is rewritten as

$$\Phi_i(n) = Im\{p_i(n)[p_i(n) + e(n)]^*\}, \quad i = 1, \ldots K. \quad (20)$$

The second order phase update equations are given by

$$\hat{\theta}_i(n+1) = \hat{\theta}_i(n) + K_{\theta_1}\Phi_i(n) + K_{\theta_2}\sum_{m=0}^{n}\Phi_i(m), \quad (21)$$
$$i = 1, \ldots K.$$

It is assumed here that the same proportional and integral tracking constants are used in all diversity channels. The update equation (21) corresponds to perfect loop integration, while it is also possible to use imperfect integration as well as sliding window integration.

The equalizer coefficients are computed adaptively based on the RLS estimation principles. The RLS algorithms have become almost a standard in digital signal processing, due to their superior convergence properties over the LMS algorithm. The RLS algorithm solves for the equalizer tap weight vector as

$$\mathbf{c}(n) = \hat{\mathbf{R}}_{uu}^{-1}(n)\hat{\mathbf{R}}_{ud}(n) \quad (22)$$

where the estimated crosscorrelation matrices are $\hat{\mathbf{R}}_{xy} = \sum_{m=0}^{n} \lambda^{n-m}\mathbf{x}(n)\mathbf{y}'(n)$, λ being the forgetting factor which accounts for the exponential windowing of the past data [11].

The long channel response (long equalizer) in conjunction with diversity reception, results in the high computational complexity of the standard RLS algorithm. A fast transversal filter (FTF) realization can be used instead for implementation. We have found a multichannel FTF algorithm presented in [12] readilly applicable for the problem at hand, with minor modifications concerning the incorporation of the carrier phases update equations.

The exact performance analysis of the proposed receiver configuration is hard to evaluate. The theoretical analysis of a similar receiver was carried out in [13] for the case of perfectly known channel responses. It is the subject of current study to evaluate the impact of estimation errors (possibly high for rapidly changing channels) on the overall receiver performance. Examining the optimal values of the receiver parameters shows that in each diversity branch, coherent combining of multipath components is performed, which results in implicit diversity improvement [13]. Further, signals from different (explicit) diversity channels are combined in a way analogous to the maximal ratio combining. In other words, if one of the channels has high SNR relative to the others, it will be favored accordingly, while if there is a 'bad' channel, with very low SNR, it will automatically be rejected in the process of adaptation.

IV. EXPERIMENTAL RESULTS

The proposed algorithm was tested and proved efficient on long range acoustic telemetry channels. The

experiment was conducted by the WHOI, off the coast of California in January 1991. The data were transmitted through the deep water over several distances ranging from 40 to 140 nautical miles and corresponding to 1, 2, 3 and 4 convergence zones. The total transmitted power was 193 $dB\ re\ 1\mu Pa$, and the data rate was varied from 3 to 333 symbols per second. The signals were received over a vertical array of 12 sensors spannning depths from $500m$ to $1500m$.

The modulation format used in the analysis is QPSK and 8-QAM. The signal is shaped at the transmitter using a cosine roll-off filter with roll of factor 0.5, and truncation length of ± 2 symbol intervals. The signals are formed using a sampling frequency of $4kHz$, and modulated onto a $1kHz$ carrier. The transmission was organized in blocks (see Fig.1). The channel probe consisted of a 13 element Barker code with rectangular (unshaped) pulses, and the data block was generated using PN sequences.

The choice of receiver parameters such as equalizer length, carrier phase tracking constants, and the forgetting factor of the RLS algorithm, was based on a series of channel estimation experiments. Here we present several examples. Fig.3 presents results of single channel reception for purposes of later comparison. It refers to QPSK transmission at 333 symbols per second over 110 nautical miles, channel 8. The channels (hydrophones of the array) are numbered $0 \div 11$, channel 0 being the one closest to the surface. Shown in the figure are the input scatter plot, the mean squared error, the phase estimate and the output scatter plot. The input SNR, measured from the Barker probe, was on the order of $14dB$. The scatter plot of the input signal is completely smeared due to the ISI, phase fluctuations and noise. The ISI at 110 nautical miles was measured to span about $60ms$, or 20 symbol intervals at the rate 333 symbols per second. The MSE indicates convergence of the algorithm in about 100 symbol intervals. The estimated phase is shown as a function of time measured in symbol intervals, after the constant Doppler shift has been removed. The output SNR, measured from the scatter plot of the estimated data symbols, was $12dB$, and the block error probability is estimated to be on the order of $2 \cdot 10^{-4}$. The used receiver parameters are indicated in the figure. The length of a feedforward equalizer is denoted by N, and the center tap was positioned in the middle.

The signal from channel 8 and the signals from channels 6 and 10 of similar characteristics and single channel performance, were processed by a multichannel algorithm, and the result is shown in Fig.4. The expected convergence period can be estimated as twice the total number of taps which is about 250 symbol intervals. The estimated phases are shown after the individual Doppler shifts of $-0.18Hz$, $-0.21Hz$, and $-0.22Hz$ were removed. The output scatter plot shows noticeable improvement in

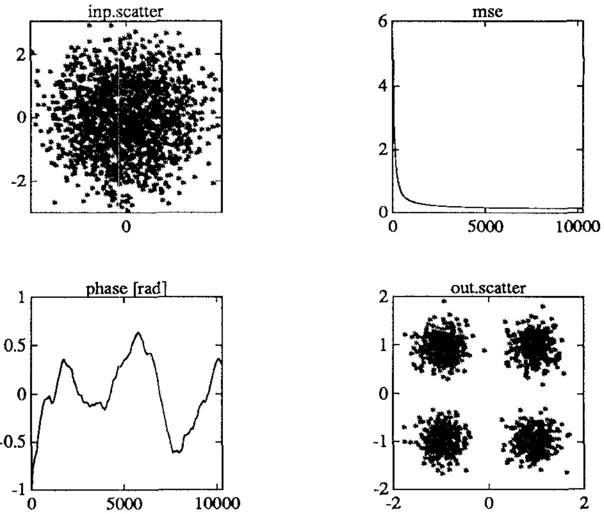

Figure 3: Results for QPSK, 333 symbols per second, 110 nautical miles, channel 8. Receiver parameters: $N = 40, M = 10, \lambda = 0.99, K_{\theta_1} = 0.001, K_{\theta_2} = 0.0001$.

performance of about $5dB$, which is to be expected for the third order diversity. There were no decision errors in this case.

Finally, Fig.5 shows multichannel results obtained with an 8-QAM signal constellation at the same rate and range. In this case, the SNR per single channel was insufficient, and multichannel processing was necessary to ensure proper operation. The output scatter plot shows successfull operation of the algorithm, with output SNR of $13.5dB$, and the probability of error on the order of 10^{-3}.

Satisfactory results were also obtained at lower rates and ranges. An attempt was made to demodulate data transmitted at 1000 symbols per second. However, the performance was limited by very poor SNR.

V. SUMMARY AND CONCLUSIONS

In the attempt to achieve reliable digital communications over long range UWA channel, we have proposed a receiver algorithm for joint, multichannel carrier phase synchronization and decision feedback equalization, based on multiparameter estimation techniques. The receiver features a second order multichannel DPLL, and an RLS adaptation of the equalizer coefficient suitable for implementation in FTF version.

The algorithm was successfully applied to the experimental long range telemetry data transmitted at rates up to 333 symbols per second, over distances covering1, 2, 3, and 4 convergence zones. The experimental results assert the fesibility of simultaneously combatting the extended ISI, and removing the phase fluctuations, thus coherently combining the multiple arrivals in each of diversity channels. Additional improvement with respect to fading and

noise, which are the major limiting factors for the UWA telemetry, is achieved through the use of spatial diversity.

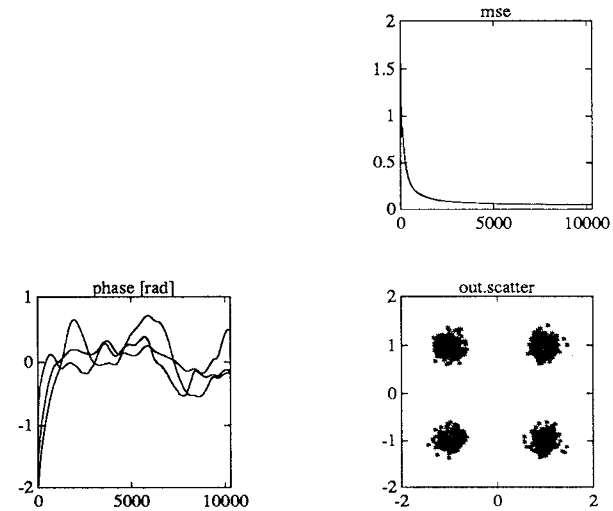

Figure 4: Results for QPSK, 333 symbols per second, 110 nautical miles, channels 6,8 and 10. Receiver parameters: $N = 40, M = 10, \lambda = 0.99, K_{\theta_1} = 0.001, K_{\theta_2} = 0.0001$.

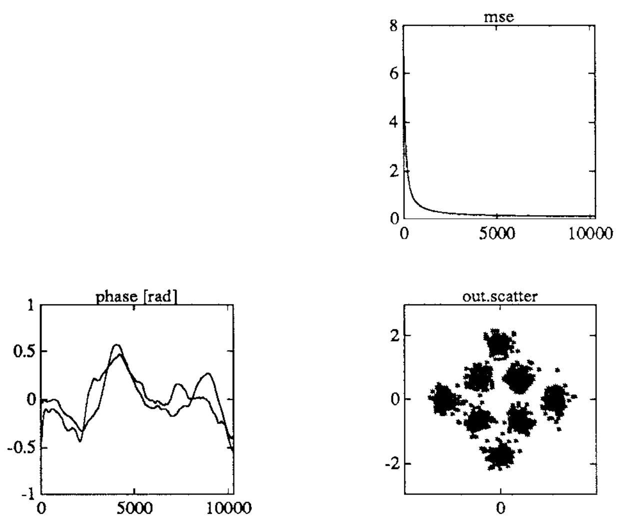

Figure 5: Results for 8-QAM, 333 symbols per second, 110 nautical miles, channels 8 and 10. Receiver parameters : $N = 40, M = 10, \lambda = 0.99, K_{\theta_1} = 0.001, K_{\theta_2} = 0.0001$.

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